# Investigation of the Equivalent Circuit Parameters and Design of a Dual Polarised Dual Frequency Aperture Coupled Microstrip Antenna

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***Abstract*: This paper presents a simplified approach to the design of a single feed dual polarised, dual frequency (1.9 and 2.4 GHz) aperture coupled microstrip antenna. Using simulation and practical investigation, the coupling between the microstrip feed line and aperture, along with the coupling between the aperture and patch, is investigated and modelled using equivalent transformers. The results obtained are used to reduce the number of interdependent design parameters, thereby allowing initial approximate values to be determined more directly with only fine tuning subsequently required to obtain good matching at both frequencies. Excellent agreement is obtained for the simulation and practical results of the return loss and the gain of the antenna.**

***Index Terms* - Aperture coupled microstrip antenna, transformer turns ratio, single feed dual polarised, dual frequency antenna.**

## I. INTRODUCTION

There is a requirement for light, low cost, wide bandwidth and high gain microwave antennas in applications such as automatic road tolling, 'on the move' satellite communication services and high-capacity data networks. These requirements can often be satisfied by an aperture coupled microstrip antenna [1], [2] where different coupling slots and stack methods have been demonstrated [3]-[5]. Matched antennas at a chosen frequency have been designed using spectral domain analysis and spatial solutions [6]-[9]. Such mathematical analysis is rigorous and elegant but because the dimensions of the physical model and the values of the circuit elements of the equivalent circuit are interdependent the design of the antenna is still difficult.

In the approach presented in this paper, practical measurements and simulation are used to determine the transmission parameter $S_{21}$, from which the turns ratio $n_f$ can be determined for an equivalent circuit transformer modelling the coupling between the feed line and the aperture. Using a curve fitting technique, a new equation for this turns ratio is proposed as a function of the length of the slot line and height of the substrate. It is found that this turns ratio is nearly independent of the dielectric permittivity of the substrate and frequency. The turns ratio $n_p$, modelling the coupling between the aperture and radiating patch, is then investigated and it is shown that the equation derived by Jaisson [10] agrees closely with obtained simulation results.

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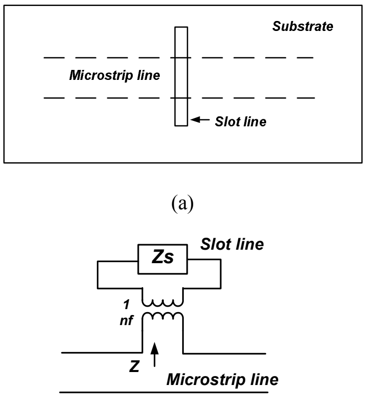
Based on these results a simplified approach is presented to design a matched dual frequency microstrip antenna.

In this work the physical antenna structures were modelled using 'Computer Simulation Technology (CST)' software while the equivalent circuit modelling employed 'Microwave Office (AWR)'software. The electrical and physical properties of the three substrates used in this work are, PCB FR4 ($\varepsilon_r$ = 4.3), Duroid 5870 ($\varepsilon_r$ = 2.3) and Duroid 6010 ($\varepsilon_r$ = 10.5) all with thickness $H$ of 1.575 mm unless otherwise specified.

## II. INVESTIGATIONS OF THE TRANSFORMERS TURNS RATIOS USED IN THE EQUIVALENT CIRCUIT OF THE ANTENNA

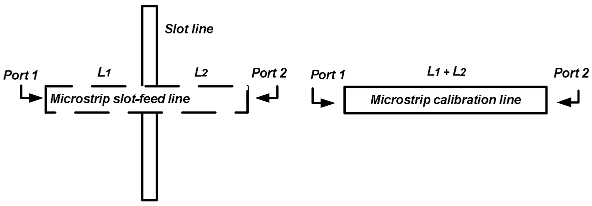
The physical structure and equivalent circuit of the dual polarised dual frequency aperture coupled microstrip antenna are well documented in [1], [2]. The physical structure of a slot line fed by a microstrip feed line is shown in Fig. 1(a) and the equivalent circuit is shown in Fig. 1(b). In the equivalent circuit the transformer (with turns ratio $n_f$) models the coupling between the slot line and the microstrip feed line, while $Z$s is the input impedance of the slot line and $Z$ is the impedance produced on the microstrip feed line [10]. For an infinitely long slot line Knorr [11], [12] and Das [13] derived equations for $n_f$ showing that it is always less than one. Bhattacharyya [14] used full wave analysis to obtain integral equations to plot this turns ratio as a function of the length of the slot line and height of the substrate. He found that this turns ratio increased with the length of the slot line and decreased with substrate thickness but did not appreciably change with the width of the slot line. In ref. [8] Bhattacharyya used spectral domain analysis to derive a large number of complex equations which had to be computed numerically to obtain a matched aperture coupled antenna. Himdi [15], Kyriacou [16] and Jaisson [10] derived different equations for $n_f$ (and similarly for $n_p$) however it was found that the equations did not agree with each other and even $n_f$ was greater than one as the length of the slot line increased. Hence in this section the turns rations $n_f$ and $n_p$ are investigated by simulation and practical measurements. To obtain the turns ratio $n_f$ it is first necessary to determine the impedance $Z$ from the $S_{21}$ parameter defined by Fig. 2(a). In this figure the slot line is fed by a 50Ω microstrip feed line of length $L_1$+$L_2$ and to calibrate out the measurements errors produced by the length $L_1$+$L_2$ a 50Ω calibration line of the same length is used in Fig. 2(b).



(a)

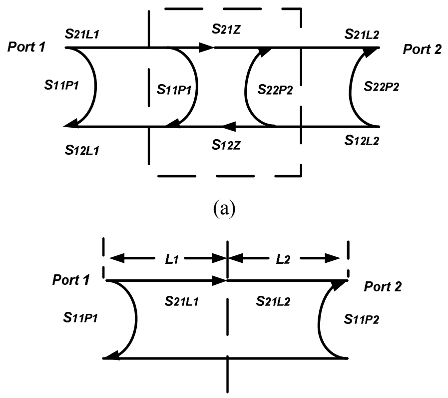
(b)

Fig. 1 (a) Microstrip-slot line structure (b) Equivalent circuit of the microstrip-slot line structure



(a) (b)

Fig. 2 (a) Signal flow graph of the feed line and the two port network (b) Signal flow graph of the calibration line



(a)

(b)

Fig. 3 Signal flow graphs (a) of the feed line and the two port network (b) of the through line

The impedance of the ports is 50Ω and the signal flow graphs of the two structures in terms of S-parameters are shown in Fig. 3. In Fig. 3(a), $S_{21L1}$, $S_{21L2}$ are the transmission parameters for the lengths $L_1$ and $L_2$ of the microstrip feed line. The impedance $Z$ is modelled by a two port network defined by $S_{11Z}$, $S_{22Z}$, $S_{21Z}$ and $S_{12Z}$. In Fig. 3(b) $S_{21L1}$ and $S_{21L2}$ are the transmission parameters of the microstrip calibration line. Equations (1) and (2) are obtained from the two signal flow graphs where, $S_{21T}$ is the transmission parameter between port 1 and port 2 in Fig. 3(a) and $S_{21C}$ between ports 1 and port 2 in Fig. 3(b).

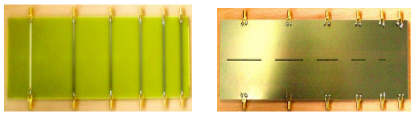


Fig. 4 Microstrip feed lines on one side, and five slot lines on the other, of a fabricated PCB board

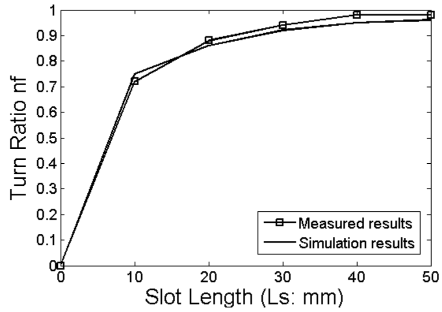



Fig. 5 Measured and simulation results for the turn ratio $n_f$ as a function slot line lengths

$$S_{21T} = S_{21L1} * S_{21Z} * S_{21L2} \quad (1)$$
$$S_{21C} = S_{21L1} * S_{21L2} \quad (2)$$

Equations (3) and (4) are used to determine $S_{21Z}$ and $Z$ by practical measurements and simulation.

$$S_{21z} = \frac{S_{21T}}{S_{21C}} \quad (3)$$

$$Z = \frac{2Z_0 - 2Z_0 S_{21z}}{S_{21z}} \quad (4)$$

The PCB FR4 board shown in Fig. 4 was used for practical measurements: one side of the board shows 5 slot lines of lengths 10, 20, 30 40 and 50mm and the other side shows the five 50Ω feed microstrip lines, as well as a 50Ω microstrip calibration line. The width of the slot line was 1mm and equations derived in [11] were used to determine the slot line input impedance $Z_S$ (see Fig. 1(b)). The turns ratio $n_f$ was obtained from the $S_{21}$paramter determined by both practical measurement and simulation. Fig. 5 shows excellent agreement between these two approaches for determining $n_f$. It is important to note that this turns ratio has a limited range of 0.9 to 0.98 for a slot line length over 25 mm that is typical in the design of an antenna at around 2 GHz.

The effect of three different substrate thicknesses (0.7875, 1.575 and 2.5mm) on $n_f$ was also investigated by simulation as described above and equation (5) was obtained from curve fitting the results.

$$n_f = 1 - e^{\frac{-Ls}{4H}} \quad (5)$$

For the three substrates of different permittivity, with a fixed slot line of length 30mm, the above method was used to obtain the frequency response of the feed line impedance $Z$ for the physical structure and the equivalent circuit. It was found that the obtained frequency response of the physical structure, even over an extended range of 1GHz to 5GHz, agreed very closely with the frequency response obtained from the equivalent circuit. As the turns ratio $n_f$ has a constant value in the equivalent circuit it can be concluded $n_f$ is effectively independent of both frequency and the relative permittivity of the substrate across the parameter ranges investigated.

In investigating the turns ratio $n_p$ modelling the coupling between the slot line and patch, first, the feed-slot dimensions were kept constant to ensure that the turns ratio $n_f$ had a fixed value and the patch width was varied from 50 mm to 80 mm. Then different slot line lengths up to 50 mm were varied and in each case $n_p$ was tuned using AWR software until the input impedance of the equivalent circuit and that of the physical structure using CST simulation were in agreement. It was found that the determined turns ratio agreed very closely with equation (6) derived by Jaisson [10] where $L_s$ is the length of the slot line and $w_p$ is physical width of the patch.

$$n_p = \frac{L_s}{2w_p} \tag{6}$$

A wide matching bandwidth and high gain antenna can be obtained by using a thick air substrate below the patch [17]. From a simulation investigation of different patch widths, equation (7) was derived for the length of the fringing field as a function of the height $H$ (mm) of the air substrate using a curve fitting approach.

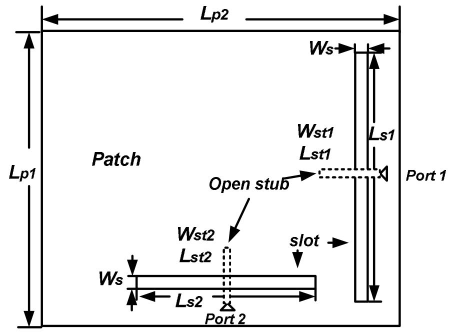
$$\Delta L = 12(1 - e^{\frac{-H}{5}}) \tag{7}$$

## III. Design of a Dual Polarised Dual Frequency Matched Antenna

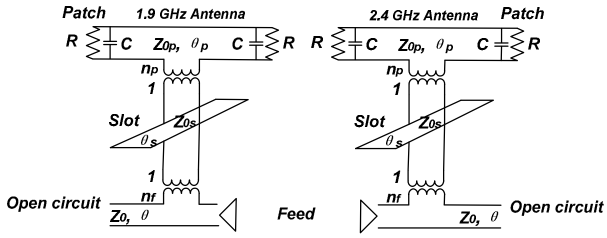
To obtain a dual frequency antenna, modes at 1.9 GHz and 2.4 GHz are excited by the two slot lines near the edges of the patch as shown in Fig. 6(a). In this figure $L_{p1}$ and $L_{p2}$ are the physical dimensions of the patch while $L_{s1}$ and $L_{s2}$ are the lengths of the two slots. The height of the air substrate above the ground plane was designed as 8 mm and the substrate below ground plane was PCB FR4. The equivalent circuits of the two modes are shown in Fig. 6(b) where $R$ models the radiated power from the patch's radiating slots and $C$ models the fringing fields at the physical edges of the patch [2].

The resonant frequency of the series input impedance of the patch at the position of slot line is normally above the design frequency. Hence the effective length of the patch should be slightly less than the half-wavelength (approx. $0.9\lambda_P/2$) where $\lambda_P$ is the patch wavelength at the design frequency. The approximate effective dimensions of the patch are 76 mm at 1.9 GHz and 61 mm at 2.4 GHz. The corresponding physical dimensions of the patch using equation (7) are 57mm at 1.9 GHz and 43mm at 2.4 GHz.

The resonant frequency of the slot line is normally higher than the design frequency so that the slot length $L_s$ should be approximately be $0.7\lambda_S/2$. Consequently the assumed initial physical lengths of the slot lines obtained are 33 mm at 2.4 GHz and 42 mm at 1.9 GHz. These lengths were then used in equation (5) and the initial approximate value of $n_f$ was 0.96 at 1.9 GHz and 0.95 at 2.4 GHz. Substituting the above slot lines lengths and the physical widths of the patch into equations (6) gives approximate values for the turns ratio $n_p$ as




(a)




(b)

Fig. 6 (a) Physical structure of the antenna (b) Equivalent circuits of the antenna at the two frequencies

0.217 at 2.4 GHz and 0.33 at 1.9 GHz. The two feed ports in Fig. 6(b) were replaced by a single feed port as shown in the photograph of the fabricated antenna in Fig. 7.

Table 1 shows that the initial and final tuned values for the antenna are close. Fig. 8 shows that a very good agreement has been obtained between the simulated and measured frequency responses for the reflection coefficient and for the gain of the antenna. Simulation results showed that gain of 6.3dB at 1.9 GHz and 7dB at 2.4 GHz with corresponding practical measurements of 6.6 dB at 1.9 GHz and 7.2 dB at 2.4GHz.

Table 1

| Dimension of dual frequency aperture coupled slot antenna | | |
|---|---|---|
| **1.9 GHz** | | |
| **Parameter** | **Initial value** | **Final tuned value** |
| ***Lp1*** | ***38 mm*** | ***43 mm*** |
| ***Lp2*** | ***53 mm*** | ***57 mm*** |
| ***Ws*** | ***1 mm*** | ***1 mm*** |
| ***Ls1*** | ***42 mm*** | ***41 mm*** |
| ***ΔLp1*** | ***9.5 mm*** | ***9.3 mm*** |
| ***Turn ratio nf*** | ***0.965*** | ***0.972*** |
| ***Turn ratio np*** | ***0.327*** | ***0.326*** |
| ***R*** | ***646 Ω*** | ***646 Ω*** |
| | | |
| **2.4GHz** | | |
| **Parameter** | **Initial value** | **Final tuned value** |
| ***Lp1*** | ***38 mm*** | ***43 mm*** |
| ***Lp2*** | ***53 mm*** | ***57 mm*** |
| ***Ws*** | ***1 mm*** | ***1 mm*** |
| ***Ls2*** | ***33 mm*** | ***35 mm*** |
| ***ΔLp2*** | ***9.5 mm*** | ***9.3 mm*** |
| ***Turn ratio nf*** | ***0.952*** | ***0.94*** |
| ***Turn ratio np*** | ***0.217*** | ***0.228*** |
| ***R*** | ***368 Ω*** | ***368 Ω*** |

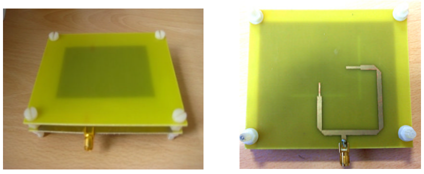
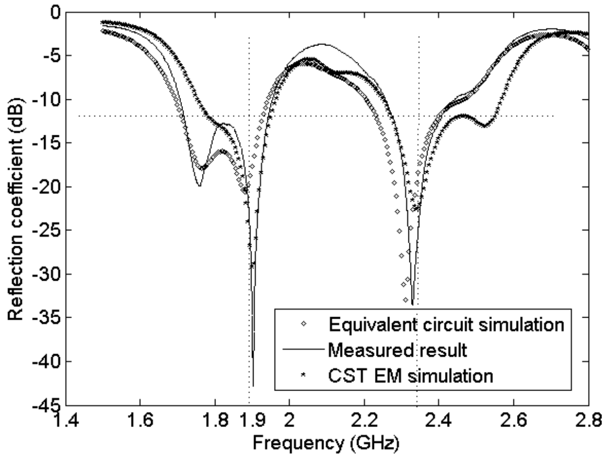
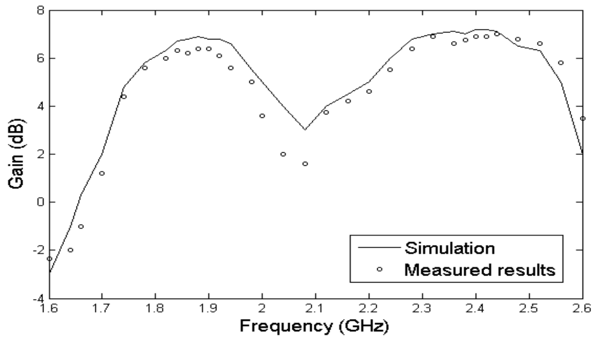


Fig. 7 Photograph of the two sides of the fabricated antenna




(a)




(b)

Fig. 8 Simulated and practical frequency response (a) Reflection Coefficient (b) Gain of the antenna

## IV. CONCLUSIONS

In this paper a simplified practical approach was used to design a dual frequency matched aperture coupled microstrip antenna. From the investigation carried out, new equations have been obtained for the turn ratios $n_f$ and the length of the fringing field at the patch. The turns ratio $n_f$ was found to have an almost constant value (0.9-0.98) for practical slot lengths of longer than 25mm used at the two design frequencies. Simulation investigation showed that the equation derived by Jaisson [10] for the turns ratio $n_p$ is sufficiently accurate to be used in the design of this antenna. Based on these results it is shown that the design complexity of the dual frequency antenna was reduced. Very good agreement was obtained between the practical and simulated results for both the frequency response of the reflection coefficient and the gain of the antenna around the two design frequencies.

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